(19) Europäisches Patentamt
European Patent Office
Office européen des brevets

(11) **EP 2 495 810 B1**

(12) **EUROPEAN PATENT SPECIFICATION**

(45) Date of publication and mention of the grant of the patent:
12.02.2014 Bulletin 2014/07

(51) Int Cl.:
H01Q 3/26 *(2006.01)* ***H04B 7/08*** *(2006.01)*
H04B 1/7075 *(2011.01)*

(21) Application number: **12168960.8**

(22) Date of filing: **24.10.2001**

(54) **PN code acquisition with adaptive antenna array**

PN-Codeerfassung mit adaptivem Antennenarray

Acquisition de code PN avec réseau d'antenne adaptative

(84) Designated Contracting States:
AT BE CH CY DE DK ES FI FR GB GR IE IT LI LU MC NL PT SE TR

(30) Priority: **24.10.2000 US 694280**

(43) Date of publication of application:
05.09.2012 Bulletin 2012/36

(62) Document number(s) of the earlier application(s) in accordance with Art. 76 EPC:
01979239.9 / 1 336 263

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US-B1- 6 201 828

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Description

BACKGROUND OF THE INVENTION

FIELD OF THE INVENTION

[0001] The present invention relates to wireless telecommunications generally, and more particularly to the application of a smart antenna to PN code acquisition in a code division multiple access (CDMA) wireless communications system.

DESCRIPTION OF THE RELATED ART

[0002] In third generation (3G) wireless communications systems, particularly wide band (W)-CDMA systems such as those described in the 3rd Generation Partnership Project, "Physical Channels and Mapping of Transport Channels onto Physical Channels (FDD)," 3GPP Technical Specification, TS25.211, v3.2.0, March 2000; 3rd Generation Partnership Project, "Spreading and Modulation (FDD)," 3GPP Technical Specification, TS25.213, v3.2.0, March 2000; and 3rd Generation Partnership Project, "FDD: Physical Layer Procedures," 3GPP Technical Specification, TS25.214, v3.2.0, March 2000 (collectively "3GPP"), or in the CDMA2000 standard in TIA, Interim V&V Text for CDMA2000 Physical Layer (Revision 8.3), March 16, 1999 ("TIA"), an option is to employ smart antenna technology in the base station. A smart antenna can suppress interfering signals from different direction of arrival angles (DOAs) from the desired users by using spatial diversity. Smart antenna technologies attract much attention these days as they support more users with a high quality of service and high data rates, up to 1.92 mega bits per second (Mbps). Examples of smart antennas and system architectures employing smart antennas may be found in commonly-owned co-pending U.S. application serial numbers 09/610,470, filed July 5, 2000, entitled "Smart Antenna with Adaptive Convergence Parameter;" published as US 6369757 09/661,155, filed September 13, 2000, entitled "Smart Antenna with No Phase Calibration for CDMA Reverse Link;" published as US 6434375 and 09/669,633, filed September 26, 2000, entitled "New Cellular Architecture." published as US 6459895.

[0003] Despite the interest in smart antenna technology generally, little attention has been paid to pseudonoise (PN) code acquisition in CDMA systems that employ a smart antenna at a base station. As used herein, PN code acquisition refers to a portion of a process referred to in the art as synchronization. Synchronization is generally regarded as encompassing two processes: PN code acquisition (in which a phase error for a known PN code is resolved to within a specified boundary - that is, a coarse PN phase code error correction), and PN code tracking, in which fine differences in PN phase code errors are detected and corrected. Thus, PN code acquisition, despite its misleading moniker, refers to a coarse correction for a PN phase error between a receiver and a transmitter (e.g., a base station and a mobile unit, or vice-versa), and does not refer to a process by which a PN code itself (as opposed to a PN code phase error) is detected.

[0004] PN code acquisition may be difficult when the smart antenna weight vector does not correspond to the desired signal's DOA (because the smart antenna will suppress signals from other DOAs). Because of this potential problem, existing systems use only one (omnidirectional) element output out of the *M* array elements for PN acquisition purposes. See F. Adachi, M. Sawahashi, and H. Suda, "Wideband DS-CDMA for Next-Generation Mobile Communications Systems," IEEE Communications Magazine, pp. 56-69, September 1998 ("Adachi et al."). Thus, the benefit of a smart antenna (e.g., the potential gain of the smart antenna) has not been used for PN code acquisition. This causes another problem as the mobile units in the system may transmit signals with low power because of the expected high smart antenna gain at a base station, with the result that the received signal-to-interference-plus-noise-ratio (SINR) at a base station may not be sufficient for PN code acquisition when only one element is employed.

BRIEF SUMMARY OF THE INVENTION

[0005] The present invention is an efficient PN code acquisition scheme which employs multiple elements of an antenna array and an adaptive threshold. The invention is particularly useful for CDMA wireless communications, especially for Direct Sequence (DS-)CDMA wireless communications. The basic structure of preferred embodiments is the combination of a conventional PN correlation searcher, an adaptive beamformer, and an adaptive threshold setting circuit. During each observation interval, which consists of multiple chips, the adaptive beamformer adaptively updates the weight vector for the smart antenna elements (as used herein, a smart antenna element refers to a single antenna, such as an omnidirectional antenna, in an array of antennas that collectively form the smart antenna) using the accumulated received signal despread with trial PN code phase error as input, preferably at the chip rate. The adaptive beam former may use any one of a number of algorithms for this purpose. In a preferred embodiment, a normalized least mean square algorithm is used. A spatially correlated signal is then formed by weighting an accumulated value of the signal received by each antenna in the array over the observation period with the corresponding final weight of smart antenna weight vector as calculated by the adaptive beamformer algorithm. This spatially correlated signal is then compared to a threshold to

determine whether PN code acquisition has occurred. If the threshold is exceeded, PN code acquisition is declared. Otherwise, a new trial PN code phase error is selected and the process is repeated.

[0006] In conventional serial search algorithms, the above-mentioned threshold for PN code acquisition has been fixed and can be calculated from a given false alarm probability P_f and a given bit-energy-to-interference power spectral density ratio E_b/I_o. But in a real mobile fading environment, E_b/I_o often varies. In preferred embodiments, an adaptive threshold setting algorithm is employed to adapt to varying environment for efficient PN code acquisition. An adaptive threshold setting algorithm has been analyzed for a receiver with a single antenna element, in Kwonhue Choi, Kyung-whoon Cheun, and Kwanggeog Lee, "Adaptive PN code Acquisition Using Instantaneous Power-Scaled Detection Threshold Under Rayleigh Fading and Gaussian Pulsed Jamming," The 4th CDMA International Conference, Proceedings Vol. II, pp. 162-169, Seoul, Korea, September 8-11, 1999 ("Choi et al."). The present invention develops an adaptive threshold setting algorithm for a receiver with multiple array elements. The adaptive threshold setting circuit is actually an average power estimator in preferred embodiments. While the adaptive beamformer updates the weight vector $\underline{w}(i)$ adaptively, the power estimator estimates the instantaneous received signal and interference power prior to PN code despreading for each observation interval of NT_c seconds. The average estimated power is then employed to scale a fixed reference threshold to create the above-mentioned adaptive threshold used to determine whether or not PN code acquisition has been achieved. The PN code acquisition time with the proposed PN code acquisition scheme with M=5 elements, by way of example, can be 210% shorter at a given SINR than the PN code acquisition time for a system, such as the system described in Adachi et al., in which only a single element is used for PN code acquisition.

[0007] In one aspect of the present invention, the PN code acquisition system may be applied to a base station, wherein the antennas are in the base station. According to another aspect of the present invention, the system is applied to a mobile station, wherein the smart antennas are in the mobile station.

BRIEF DESCRIPTION OF THE DRAWINGS

[0008] The features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIGURE 1 shows a block diagram of a proposed PN code acquisition scheme for a DS-CDMA system configured in accordance with one embodiment of the present invention;

FIGURE 2 shows a flow chart of the proposed PN code acquisition scheme for a DS-CDMA system configured in accordance with one embodiment of the present invention;

FIGURE 3 shows the theoretical and simulation false alarm probabilities P_f versus bit-energy-to-interference power spectral density ratio E_b/I_o in dB for M=1, 3, and 5 elements cases under a fading environment, indicating that the PN code acquisition scheme configured in accordance with one embodiment of the present invention is robust against the received signal power variations;

FIGURE 4 shows the theoretical and simulation detection probabilities P_d versus bit-energy-to-interference power spectral density ratio E_b/I_o in dB for M=1, 3, and 5 elements under fading environment, indicating that the PN code acquisition scheme configured in accordance with one embodiment of the present invention improves detection probability P_d significantly; and

FIGURE 5 shows the theoretical and simulation average acquisition time T_{acq} versus bit-energy-to-interference power spectral density ratio E_b/I_o in dB for M=1, 3, and 5 elements under fading environment, indicating that the PN code acquisition scheme configured in accordance with one embodiment of the present invention improves average acquisition time T_{acq} significantly.

DETAILED DESCRIPTION

[0009] In the following detailed description, many specific details, such as chip rates and numbers of elements in smart antenna arrays, are provided by way of example in order to provide a thorough understanding of the present invention. The details discussed in connection with the preferred embodiments should not be understood to limit the present invention.

[0010] The present invention is described below in the context of a reverse link from a mobile to a base station in a CDMA wireless communications system. However, the invention is equally applicable to forward link PN code acquisition. The invention is also believed to be particularly applicable to Direct Sequence CDMA (DS-CDMA) systems, and thus will be discussed in the context of the same herein, but should not be understood to be limited to DS-CDMA systems.

[0011] FIGURE 1 shows a block diagram of the proposed PN code acquisition scheme for a DS-CDMA system configured in accordance with one embodiment of the present invention. The input signal $r'(t)$ received from the each element 101 of the smart antenna array is passed through a demodulator 102 and a matched filter 103. The output of

each matched filter 103 is sampled at the chip rate by a chip sampler 104, and correlated (despread) with the trial PN code phase error by a correlator 105. The output of the correlators 105 is stored by accumulators 106. The output of the accumulators 106 is used by an update process for the smart antenna weight vector.

[0012] In preferred embodiments, the weight vector of the smart antenna is updated with an LMS algorithm. In highly preferred embodiments, the normalized least mean square (LMS) algorithm described in S. Haykin, Adaptive filter Theory, 3rd Edition, Prentice Hall, Upper Saddle River, New Jersey, 1996 ("Haykin"), is employed to update the weight vector. Other vector update algorithms are also possible, including those algorithms discussed in commonly assigned co-pending U.S. application 09/610,470, filed July 5, 2000, entitled "Smart Antenna with Adaptive Convergence Parameter." It is not necessary for the vector update to be an LMS algorithm. During each observation interval of NT_c (where T_c is the chip period and N is the number of chip periods in an observation interval) seconds, an adaptive beamformer 107 updates the weight vector $\underline{w}(i)$ adaptively, preferably at the chip rate. The input to adaptive beamformer 107 is the accumulator 106 output after despreading using the trial PN code phase error as described above.

[0013] In preferred embodiments, the vector update process starts after the observation interval begins. This is done in order to conserve processing power. The sample time index $i=N_s$ (where i is the chip or sample time index) on which the update starts is chosen to reflect the time required for the adaptive algorithm to converge. For example, in preferred embodiments using the normalized LMS algorithm described above and in which the number of chips N in the observation period is equal to 256, N_s is chosen as 192, which provides for 256-192=64 iterations of the algorithm. This reflects that fact that the algorithm typically takes approximately 40 iterations to converge and includes a safety factor.

[0014] The update process ends at *i=N-2* for the first observation interval. The last weight vector $\underline{w}(N-1)$ calculated by the update algorithm in an observation interval, which can be regarded as a best weight vector $\underline{w}_{opt}$, is used to generate an antenna array spatial correlation output $\bar{z}$ with the last accumulation output at *i=N-1*. The square of $\bar{z}$ is then compared to an adaptive threshold (discussed in further detail below) to determine whether PN code acquisition has been achieved. If the threshold is exceeded, the PN code has been acquired and the process ends. Otherwise, a new trial PN code phase error is selected and the algorithm is repeated. The weight vector adaptation is performed at each successive observation interval of NT_c seconds until PN code acquisition is achieved.

[0015] The preferred LMS algorithm takes only *5M* computations per chip period, where *M* is the number of the antenna array elements 101. The update rate can be sufficiently small such as a chip rate of *1.2288 Mcps.* The present invention employs preferably all array elements 101 of an array for a PN correlation searcher while only one element is employed in the existing literatures ("Adachi et al.").

[0016] The following discussion explains the above-described process in greater detail. The array antenna elements 101 are assumed to be substantially identical and have the same response to any direction and the antenna spacing is one half of the carrier wavelength. The array response vector can be written as

$$\underline{a}_0(\theta) = [1 \quad e^{-j\pi\sin\theta} \quad \ldots\ldots \quad e^{-j\pi(M-1)\sin\theta}]^T \qquad (1)$$

where θ is the DOA from the desired user signal, and T denotes the transpose. The received signal at the m-th element can be written as

$$r'_m(t) = \sqrt{2P}\alpha(t)b(t-\tau T_c)c(t-\tau T_c)e^{j(w_c t+\varphi_m)}e^{j\phi(t)}e^{-j\pi(m-1)\sin\theta} + n_{m,BPF}(t) \qquad (2)$$

where *P* is the received signal power, $\alpha(t)$ and $\phi(t)$ are the fading amplitude and phase, respectively, b(t) is information data, *c(t)* is the PN spreading sequence, w_c and φ_m are the angular carrier frequency and phase, respectively, T_c is a chip interval, and τ is a PN code phase offset from a reference, which is assumed to be a random integer uniformly distributed between *0* and PN sequence period *L*. The $n_{m,BPF}(t)$ in equation (2) represents an additive white Gaussian noise (AWGN) bandpass interference plus noise signal. A fading channel with a single path is assumed for simplicity. The present invention can be applied for a multipath channel also. A simplified search, which shifts the trial PN code sequence by T_c, is considered here for simplicity. The phase of the PN code can be shifted in a fraction of chip interval in a complete search and it takes long simulation time. A single dwell is assumed for simplicity. The invention can be applied for a complete and multi-dwell PN code searches also. The PN spreading sequence c(t) is written as

$$c(t) = \sum_i c(i) \prod_{T_c} (t - iT_c) \tag{3}$$

where c(i) $\in$*{1, -1}* with equal probability, Π_{Tc} is a rectangular shape pulse with unit amplitude and duration T_c. The carrier synchronization is not assumed because noncoherent PN code single dwell serial search is employed. Also, assume a pilot channel is used instead of pilot symbols, i.e., *b(t)=1* for all time as the pilot channel in cdma2000. Then, the equivalents baseband signal $r_m(t)$ at the m-th element after down-conversion with carrier frequency can be written as

$$r_m(t) = \sqrt{P}\alpha(t)c(t - \tau T_c)e^{j[\varphi_m + \phi(t) - \pi(m-1)\sin\theta]} + n_m(t) \tag{4}$$

where $n_m(t)=n_m^I(t)+jn_m^Q(t)$ is a complex AWGN with mean zero and variance I_0. The output of matched filter $H^*(f)$ 103 is sampled at chip rate and the samples $r_m(i)$ are written as

$$r_m(i) = \sqrt{E_c}\alpha(i)\, c(i - \tau)e^{j[\varphi_m + \phi(i) - \pi(m-1)\sin\theta]} + n_m(i) \tag{5}$$

where i and E_c denote the chip index and chip energy, respectively. After correlating with the local PN code at 105, $r_m(i)$ becomes

$$x_m(i) = [\sqrt{E_c}\alpha(i)\, c(i - \tau)e^{j\phi_m(i)} + n_m(i)]c(i - \hat{\tau}) \tag{6}$$

where $\hat{\tau}$ is the estimated PN code phase and

$$\phi_m(i) = \varphi_m + \phi(i) - \pi(m-1)\sin\theta\,. \tag{7}$$

[0017] All antenna element outputs can be written in a vector form as

$$\underline{x}(i) = [\sqrt{E_c}\alpha(i)\, c(i - \tau)\underline{a}(\phi) + \underline{n}(i)]c(i - \hat{\tau}) \tag{8}$$

where

$$\underline{a}(\phi) = [e^{j\phi_1(i)} \quad e^{j\phi_2(i)} \quad \ldots \quad e^{j\phi_M(i)}]^T \tag{9}$$

can be regarded as a modified array response vector, compared with $\underline{a}_0(\theta)$ in equation (1). Assume an observation interval of NT_c for a PN code phase shift. Energy is accumulated every chip at each branch after PN despreading 105. The accumulator 106 output at iT_c for the m-th branch can be written as

$$y_m(i) = \sum_{i'=0}^{i} x_m(i') \tag{10}$$

where *i=0,* 1, ..., *N-1* for the first observation interval, and *m=1, 2, ..., M.* The adaptive beamformer 107 preferably starts

updating weight vector $\underline{w}(i)$ adaptively at $i=N_s$ and ends updating at $i=N-2$ for the first observation interval. (Those of skill in the art will recognize that it is also possible to begin updating earlier or later, even as early as the start of the observation interval.) The same process is applied for the other observation intervals until PN code is acquired. The normalized LMS algorithm is employed to update the weight vector. This weight vector adaptation is performed every observation interval of NT_c seconds. The input to the adaptive beamformer 107 is written as

$$\underline{y}(i) = [y_1(i) \quad y_2(i) \quad \ldots \quad y_M(i)]^T. \tag{11}$$

[0018] The last weight vector $\underline{w}(N-1)$ in an observation interval, which can be regarded as a best weight vector $\underline{w}_{opt}$, is used to generate a spatial correlation output $\bar{z}$ with the last accumulation output $\underline{y}(N-1)$. The antenna array spatial correlation output $\bar{z}$ 109 can be written as

$$\begin{aligned}\bar{z} &= \underline{w}^H(N-1)\underline{y}(N-1)/(MN) \\ &= \frac{1}{MN}\sum_{m=1}^{M}\sum_{i=0}^{N-1}[w_{optm}\, x_m(i)] \\ &= \frac{1}{MN}\sum_{j=1}^{MN} z_j\,,\end{aligned} \tag{12}$$

where H stands for Hermitian, i.e., conjugate and transpose,

$$\underline{w}(N-1) = \underline{w}_{opt} = [w_{opt1} \quad w_{opt2} \quad \ldots \quad w_{optM}]^H, \tag{13}$$

and

$$z_j = w_{optm}\, x_m(i), \qquad j = (m-1)N + i + 1. \tag{14}$$

[0019] The final decision variable Z 110 is normalized and written as

$$Z = |\bar{z}|^2 = \left|\frac{1}{MN}\sum_{j=1}^{MN} z_j\right|^2. \tag{15}$$

[0020] As discussed above, this final decision variable Z is then compared to a threshold, preferably an adaptive threshold as calculated by an adaptive threshold setting circuit 108, in order to determine whether PN code acquisition has been achieved. In preferred embodiments, the adaptive threshold setting circuit 108 estimates the average power during each observation interval NT_c. Referring to FIGURE 1 and considering $|r_m(i)|2=|xm(i)|^2$, we can get the average power $\bar{P}$ as

$$\overline{P} = \sum_{m=1}^{M} \{[\frac{1}{M}|w_{optm}|^2] \cdot [\frac{1}{N}\sum_{i=0}^{N-1}|r_m(i)|^2]\}$$

$$= \frac{1}{MN}\sum_{m=1}^{M}\sum_{i=0}^{N-1}|w_{optm}\, x_m(i)|^2$$

$$= \frac{1}{MN}\sum_{j=1}^{MN}|z_j|^2 . \qquad (16)$$

[0021] If Z is larger than a scaled threshold $k \cdot \overline{P}$, a PN code acquisition is declared and the tracking loop is triggered for a single dwell serial search, where k is the fixed reference threshold. Otherwise, the acquisition scheme shifts the locally generated PN code phase (i.e., creates a new trial PN code phase error), and the searcher continues until a correct PN code phase is claimed (i.e., PN code acquisition is achieved).

[0022] The probability of Z being greater than $k \cdot \overline{P}$ is given as

$$P_r\{Z > k \cdot \overline{P}\} = P_r\{\left|\frac{1}{MN}\sum_{j=1}^{MN} z_j\right|^2 > \frac{k}{MN}\sum_{j=1}^{MN}|z_j|^2\} . \qquad (17)$$

[0023] To ease the derivation of the false alarm probability P_f and detection probability P_d later, we rewrite equation (17) using new decision variables U and V as

$$P_r\{Z > k \cdot \overline{P}\} = P_r\{U > V\} \qquad (18)$$

with

$$U = (1-k)MN|\overline{z}|^2 , \qquad (19)$$

$$V = k(MN-1)\overline{v}_z , \qquad (20)$$

where $\overline{z}$ is the sample mean of z_j given in equation (12), and $\overline{v_z}$ denotes the sample variance of z_j given as

$$\overline{v}_z = \frac{1}{MN-1}\sum_{j=1}^{MN}|z_j - \overline{z}|^2 . \qquad (21)$$

[0024] Under hypothesis H_1 when the received PN code and the local PN code are aligned with the same phase, equation (18) gives the detection probability P_d, and under hypothesis H_0 when PN codes are out of phase, equation (18) gives the false alarm probability P_f as follows:

$$P_d = P_r\{U > V|H_1\} , \qquad (22)$$

$$P_f = P_r\{U > V | H_0\}. \quad (23)$$

[0025] From equations (19) and (20), we observe that both P_d and P_f are equal to 1 when k<0 and both equal to 0 when k>1. So we focus on the range of *k* between 0 and *1* in this invention.

[0026] The following is the theoretical analysis. Assume a slow Rayleigh fading channel so that the amplitude $\alpha(i)$ as well as the phase ϕ(i) remain constant over each observation time NT_c but are independent between the observation intervals. Then $\alpha(i)$ and $\phi(i)$ can be written as

$$\alpha(i) = \alpha, \qquad 0 \le i \le N-1, \quad (24)$$

$$\phi(i) = \phi, \qquad 0 \le i \le N-1, \quad (25)$$

for the first observation interval, where α and ϕ are constants. Therefore, (7) becomes

$$\phi_m(i) = \varphi_m + \phi - \pi(m-1)\sin\theta = \phi_m\,, \quad 0 \le i \le N-1, \quad (26)$$

for the first observation interval, where ϕ_m is constant.

(1) When PN code is synchronized

[0027] This case means

$$c(i-\tau) = c(i-\hat{\tau}). \quad (27)$$

[0028] For a theoretical analysis, we apply perfect weight vector $\underline{w}_{pft}$ equal to the modified array vector in equation (9) to obtain an ideal performance when PN code is synchronized. In other words, we assume that the smart antenna can track the DOA of the desired user signal perfectly as

$$\underline{w}_{pft} = \underline{a}(\phi) \quad (28)$$

during the end period of an observation interval, $N_s \le i \le N-1$, when PN code is synchronized. In practice, a smart antenna cannot track perfectly the DOA of the desired user. The ideal case in equation (28) provides an upper and a lower bound of detection probability P_d and average acquisition time T_{acq}, respectively.

[0029] Equation (28) implies

$$w_{opt,m} = w^*_{pft,m} = e^{-j\phi_m}\,, \qquad 1 \le m \le M. \quad (29)$$

[0030] Substituting equations (6), (27) and (29) into (14), we have

$$z_j = \sqrt{E_c}\,\alpha + n_j\,, \qquad j = (m-1)N + i + 1, \quad (30)$$

where $n_j=e^{-j\phi_m}n_m(i)c(i-\hat{\tau})$, which can be easily shown to be a complex AWGN variable with zero mean and variance equal to I_0. Substituting equation (30) into equation (12), we get the sample mean

$$\bar{z} = \sqrt{E_c}\,\alpha + \bar{n}, \tag{31}$$

where $\bar{n} = \frac{1}{MN}\sum_{j=1}^{MN} n_j$ is a complex AWGN variable with zero mean and variance equal to I_0/MN. Substituting equations (30) and (31) into equations (19) and (20), we have

$$U = (1-k)MN\left|\sqrt{E_c}\,\alpha + \bar{n}\right|^2, \tag{32}$$

$$V = k(MN-1)\bar{v}_n, \tag{33}$$

where $\bar{v}_n = \frac{1}{MN-1}\sum_{j=1}^{MN}\left|n_j - \bar{n}\right|$, is the sample variance of n_j. From equation (32),

we can show the decision variable U has the non-central chi-square distribution with its conditional probability density function as

$$p_{U1}(u|\alpha) = \frac{\exp\{-[u+(1-k)MNE_c\alpha^2]/[(1-k)I_0]\}}{(1-k)I_0} I_0\left(\frac{2\sqrt{(1-k)MNE_c\alpha^2 u}}{(1-k)I_0}\right), u \geq 0, \tag{34}$$

where $I_0(x)$ is the zero-th order modified Bessel function of the first kind, and α is Rayleigh-distributed with its probability density function as

$$p(\alpha) = \frac{2\alpha}{\sigma_f^2} e^{-\alpha^2/\sigma_f^2} \tag{35}$$

where σ_f^2 is the average fading channel power. The probability density function (pdf) of U under synchronization is

$$p_{U1}(u) = \int_0^\infty p_{U1}(u|\alpha)\,p(\alpha)\,d\alpha = \frac{\exp(-u/\beta)}{\beta}, \quad u \geq 0, \tag{36}$$

where

$$\beta = (1-k)(MNE_c\sigma_f^2 + I_0). \tag{37}$$

[0031] From equation (33), we can show the decision variable V has the central chi-square distribution with *2(MN-1)*

degrees of freedom with its probability density function as below :

$$p_{V1}(v) = \frac{v^{(MN-2)} \exp[-v/(2\sigma^2)]}{\sigma^{2(MN-1)} 2^{(MN-1)} \Gamma(MN-1)} \quad , \qquad v \geq 0, \tag{38}$$

where

$$\sigma^2 = kI_0/2 \, . \tag{39}$$

[0032] From equations (32) and (33), we know *U* and *V* are statistically independent because sample mean and sample variance are statistically independent when sampled from a Gaussian distribution. Then the detection probability can be derived to be

$$P_d = P_r\{U > V | H_1\} = [\frac{\beta}{\beta + 2\sigma^2}]^{MN-1}$$

$$= [\frac{(1-k)(MNE_c\sigma_f^2 + I_0)}{(1-k)MNE_c\sigma_f^2 + I_0}]^{MN-1} \, . \tag{40}$$

(2) When PN code is not synchronized

[0033] This case means

$$c(i-\tau) \neq c(i-\hat{\tau}). \tag{41}$$

[0034] Assume the beamforming direction of smart antenna is different from the DOA of the desired user signal. This is true in general when the PN code is not synchronized. Assume that the weight vector can be written as

$$\underline{w}_{opt} = \underline{w}(N-1) = \underline{a}(\eta) = [e^{j\eta_1} \quad e^{j\eta_2} \quad \ldots \quad e^{j\eta_M}]^T \neq \underline{a}(\phi) \tag{42}$$

without loss of generality. Equation (40) implies

$$w_{optm} = e^{-j\eta_m} \neq e^{-j\phi_m} \, , \qquad 1 \leq m \leq M. \tag{43}$$

[0035] Substituting equations (6), (41) and (43) into equation (14), we have

$$z_j = \sqrt{E_c}\, \alpha\, c(i-\tau)\, c(i-\hat{\tau}) e^{j(\phi_m - \eta_m)} + e^{-j\eta_m} n_m(i)\, c(i-\hat{\tau}). \tag{44}$$

[0036] We can show z_j is a complex AWGN variable n_0 with zero mean and variance equal to $(E_c\sigma_f^2+I_0)$. Equation (44) can be rewritten as

$$z_j = n_0. \tag{45}$$

[0037] Substituting equation (45) into equation (12), we have

$$\bar{z} = \frac{1}{MN}\sum_{j=1}^{MN} n_0 = \bar{n}_0, \tag{46}$$

which is also a complex AWGN variable with zero mean and variance equal to $(E_c\sigma_f^2+I_0)/(MN)$. Substituting equations (45) and (46) into equations (19) and (20), we have

$$U = (1-k)MN\left|\bar{n}_0\right|^2, \tag{47}$$

$$V = k(MN-1)\bar{v}_{n_0}, \tag{48}$$

where $\bar{v}_{n_0} = \frac{1}{MN-1}\sum_{j=1}^{MN}\left|n_0-\bar{n}_0\right|$ is the sample variance of n_0. Similarly as subsection (1), we can show the decision variables U and V follow central chi-square distribution with 2 and *2(MN-1)* degrees of freedom, respectively, and their probability density function are given as below

$$p_{U0}(u) = \frac{\exp(-u/\beta)}{\beta}, \quad u \geq 0, \tag{49}$$

$$p_{V0}(v) = \frac{v^{(MN-2)}\exp[-v/(2\sigma^2)]}{\sigma^{2(MN-1)}2^{(MN-1)}\Gamma(MN-1)}, \quad v \geq 0, \tag{50}$$

where

$$\beta = (1-k)(E_c\sigma_f^2+I_0), \tag{51}$$

$$\sigma^2 = k(E_c\sigma_f^2+I_0)/2. \tag{52}$$

[0038] Similarly as subsection *(1),* we also know *U* and *V* are statistically independent from equations (47) and (48),

due to the fact that sample mean and sample variance are statistically independent when sampled from a Gaussian distribution. Then the false alarm probability can be derived to be

$$P_f = P_r\{U > V | H_0\} = [\frac{\beta}{\beta + 2\sigma^2}]^{MN-1}$$

$$= (1-k)^{MN-1} . \quad (53)$$

[0039] We observe that the false alarm probability P_f is a function of *k, M* and *N* only and does not depend on E_c and I_0. Thus the adaptive threshold algorithm can approximately achieve a constant false alarm rate (CFAR).

[0040] FIGURE 2 shows the corresponding flow chart of the proposed PN code acquisition system with the present invention. All the vectors in FIGURE 2 have *M* elements, where *M* is the number of antenna elements. First, we start from the received sequence $\underline{r}(i)$ in block 201, which is formed by the output of the chip samplers 104 in FIGURE 1, where *i* is the time index. Next, $\underline{x}(i)$ (the despread signal) in block 202 is formed by the output of the correlators 105. Next, the $\underline{y}(i)$ in block 203 is formed from the accumulator 106 output. The output of block 203 is then used in block 204 and block 208. At $i=N_s$ the adaptive beamformer 107 starts updating the weight vector $\underline{w}(i)$. $\underline{W}_0=(1\ 1\ ...\ 1)^H$ in block 205 is the initial weight vector at $i=N_s$ In block 204, the input to the adaptive beamformer 107 is $\underline{\hat{y}}(i)$, which is the normalized $\underline{y}(i)$. The output of block 204 is used in updating the weight vector block 206. Block 206 shows the LMS algorithm for the adaptive beamformer 107. As shown in block 206, an intermediate value $\underline{w}(i)$ of the weight vector is formed at each update, which is then used at the subsequent update. The updated weight vector from block 206 is output to block 207. At *i=N-1,* the final weight vector shown in block 207 is multiplied with the final accumulator output $\underline{y}(N-1)$ shown in block 208 to generate the spatial correlation output 109 $\bar{z}$ in block 209. Block 210 calculates the magnitude square of $\bar{z}$ using the output of block 209 to get the decision variable Z 110, which is output to blocks 212. The power is estimated at block 211 utilizing the outputs of block 201 and 207. In block 212, Z is compared with the threshold scaled by the adaptive threshold setting circuit 108 output shown in block 211 and then the decision as to whether to declare PN code acquisition is made.

[0041] To verify the theoretical results, simulation has been done with the PN code acquisition scheme configured in accordance with one embodiment of the present invention. The simulation parameters are described as the following. P_f*=0.01,* the number of antenna elements *M =1,* 3, and *5,* N_s*=192, N=256,* bit rate *=9600 bits/second,* chip rate $=1.2288\times10^6$ *chips/second,* and penalty factor K_p*=256* were assumed. The fixed reference threshold *k* was chosen to guarantee P_f *to* be equal to *0.01* for the adaptive threshold case. *k=0.0179* was chosen for one element case; *k=0.0108* for three element case; and *k=0.0089* for five elements case. The fixed reference I_0 was chosen to make P_f in equation (59) to be equal to *0.01* at E_b/I_o*=-15 dB* for the nonadaptive threshold case. One sample per chip was taken for simulation.

Jakes' fading model was used for fading environment. The average fading channel power $\sigma_f^2=1$, velocity v*=80 miles/hour* and carrier frequency f_c *= 900 M*Hz were chosen for fading case. Spreading factor *128* and $E_b/I_o = 128E_c/I_o$ were used where E_b is information bit energy. The simulation parameters are summarized in Table 1.

Table 1

Simulation Parameters	
Description	**Parameter Value**
False Alarm Probability	$P_f = 0.01$
Observation Interval	$N = 256$
Starting Time Index of Updating Weight Vector	$N_s = 192$
Number of Samples per Chip	*1 sample/chip*
Penalty Factor	$K_p = 256$
Bit Rate	*9600 bits/second*
Chip Rate	1.2288×10^6 *chips/second*

(continued)

Simulation Parameters		
Description		**Parameter Value**
Spreading Factor		*128*
Bit-Energy-to-Interference Power Spectral Density Ratio		$E_b/I_o = 128\ E_c/I_o$
Carrier Frequency		$f_c = 900$ *M*Hz
Mobile User Velocity		*v= 80 miles/hour*
Average Fading Channel Power		$\sigma_f^2 = 1$
Fixed Reference Threshold k^1	*M = 1*	*k= 0.0179*
	M = 3	*k= 0.0108*
	M = 5	*k= 0. 089*
Interference Power Spectral Density Ratio	*M = 1*	$I_0 = 0.9947$
	M = 3	$I_0 = 0.9981$
	M = 5	$I_0 = 0.9987$
[1] *k* is chosen such that the PN code acquisition yield a given P_f=0.01 for the adaptive threshold case. When $\overline{P}$ =1, *k* becomes the fixed threshold for the nonadaptive threshold case. That is why *k* is called the fixed reference threshold.		

[0042] FIGURE 3 shows the theoretical and simulation false alarm probabilities P_f versus bit-energy-to-interference power spectral density ratio E_b/I_o in dB for *M*=1, 3, and 5 cases under fading environment with a PN code acquisition scheme configured in accordance with one embodiment of the present invention. The simulation P_f for *M*=1, 3, and 5 cases fluctuate very slightly around the theoretical P_f*=0.*01. And the simulation P_f*'s* do not increase as E_b/I_o increases while P_f does for a nonadaptive threshold scheme. This means the present invention can approximately achieve CFAR and is robust against the received signal power variations.

[0043] FIGURE 4 shows the theoretical and simulation detection probabilities P_d versus bit-energy-to-interference power spectral density ratio E_b/I_o in dB for *M*=1, 3, and 5 under fading environment, with a PN code acquisition scheme configured in accordance with one embodiment of the present invention. At P_d*=0.8, M=3* and 5 case simulation results show *3.5 dB* and *4.9 dB* improvement in SINR, respectively, compared with *M=1* case.

[0044] FIGURE 5 shows the theoretical and simulation average acquisition time T_{acq} versus bit-energy-to-interference power spectral density ratio E_b/I_o in dB for *M=1,* 3, and 5 under fading environment with a PN code acquisition scheme configured in accordance with one embodiment of the present invention. At T_{acq}*=10*^{4}*NT*$_c$*, M=3* and 5 case can improve *2.7 dB* and *4.2 dB* in SINR, respectively, compared with *M*=1 case. Therefore, when a base station employs the present invention, a random access user may transmit signal with smaller power than the single antenna case. Interference can be reduced and capacity of the system can be increased. Or at E_b/I_o*=-10 dB,* the PN code acquisition time with the proposed PN code acquisition scheme of *M*=5 elements can be significantly shortened, e.g., *210%,* compared to the existing PN code acquisition scheme of a single element in "Adachi et al." We observe that simulation results are different from the theoretical results when *M*≥2. This is because the smart antenna weight vector does not perfectly match with the array response vector of the desired user signal's DOA while perfect weight vector was assumed for the theoretical results.

[0045] In conclusion, a PN code acquisition scheme using a smart antenna and an adaptive threshold setting of the present invention is robust against the received signal power variations and could improve performance *210%* in PN code acquisition time or at least *4.5* dB in SINR to achieve the same detection probability when *M*=5 elements are used.

[0046] Those of skill in the art will readily appreciate that various alternative designs and embodiments for practicing the invention are possible. For example, while a single dwell has been described above, the present invention is equally applicable to other PN searches, e.g., double dwell, serial and complete searches. Additionally, those of skill in the art will recognize that the various devices described herein (e.g., accumulators) can be implemented in hardware and/or software. It is therefore to be understood that within the scope of the appended claims, the invention may be practiced otherwise than as specifically described herein.

Claims

1. A method for calculating a threshold used to determine whether pseudo-noise (PN) code acquisition has occurred, the method comprising the steps of:

   accumulating received signals that have not been despread from each element of a plurality of received paths to form a plurality of separately maintained unspread accumulated signals;
   weighting each unspread accumulated signal from each received path by a corresponding weight from a weight vector; and
   combining the weighted unspread accumulated signals.

2. A method according to claim 1, further comprising the step of averaging the combined weighted unspread accumulated signals.

3. A method according to claim 2, further comprising the step of repeating the accumulating, weighting, combining and averaging steps during each observation period in which the threshold is used.

4. An apparatus for calculating a threshold used to determine whether pseudo-noise (PN) code acquisition has occurred, the apparatus comprising:

   a plurality of second accumulators, each of the second accumulators being connected to receive a square of an output from a corresponding chip sampler, the chip sampler being connected to receive a signal from a corresponding element of a plurality of received paths;
   a plurality of multipliers, each multiplier being connected to a respective accumulator to receive a final accumulated value, each multiplier further being configured to weight the final accumulated value by a weight based on a corresponding weight from a weight vector; and
   an adder, the adder being connected to each of the multipliers to form a sum of outputs from the multipliers.

5. An apparatus according to claim 4, further comprising means for averaging the combined weighted unspread accumulated signals.

6. An apparatus according to claim 5, further comprising means for repeating the accumulating, weighting, combining and averaging steps during each observation period in which the threshold is used.

Patentansprüche

1. Verfahren zum Berechnen eines Schwellenwerts, der verwendet wird, um zu bestimmen, ob es zu einer Pseudozufallsfolgen (PN)-Erfassung gekommen ist, wobei das Verfahren die folgenden Schritte umfasst:

   Akkumulieren von empfangenen Signalen, die nicht entspreizt wurden, von jedem Element einer Mehrzahl von empfangenen Pfaden, um eine Mehrzahl von separat gehaltenen ungespreizten akkumulierten Signalen zu bilden;
   Gewichten jedes ungespreizten akkumulierten Signals von jedem empfangenen Pfad durch ein entsprechendes Gewicht von einem Gewichtungsfaktor; und
   Kombinieren der gewichteten, ungespreizten akkumulierten Signale.

2. Verfahren nach Anspruch 1, ferner den Schritt der Bildung eines Durchschnitts aus den kombinierten, gewichteten ungespreizten akkumulierten Signalen beinhaltend.

3. Verfahren nach Anspruch 2, ferner den Schritt des Wiederholens der Akkumulierungs-, Gewichtungs-, Kombinierungs- und Durchschnittsbildungsschritte während jedes Beobachtungszeitraums, in dem der Schwellenwert verwendet wird, beinhaltend.

4. Vorrichtung zum Berechnen eines Schwellenwerts, der verwendet wird, um zu bestimmen, ob es zu einer Pseudozufallsfolgen (PN)-Erfassung gekommen ist, wobei die Vorrichtung aufweist:

   eine Mehrzahl von zweiten Akkumulatoren, wobei jeder von den zweiten Akkumulatoren so angeschlossen ist,

dass er ein Quadrat eines Ausgangswerts von einem entsprechenden Chip-Sampler empfängt, wobei der Chip-Sampler so angeschlossen ist, dass er ein Signal von einem entsprechenden Element einer Mehrzahl von empfangenen Pfaden empfängt;
eine Mehrzahl von Multiplizierern, wobei jeder Multiplizierer mit einem entsprechenden Akkumulator verbunden ist, um einen endgültigen akkumulierten Wert zu empfangen, wobei jeder Multiplizierer ferner so ausgelegt ist, dass er den endgültigen akkumulierten Wert mit einem Gewicht auf Basis eines entsprechenden Gewichts von einem Gewichtungsfaktor gewichtet; und
einen Addierer, wobei der Addierer mit jedem von den Multiplizierern verbunden ist, um eine Summe aus Ausgangswerten von den Multiplizierern zu bilden.

5. Vorrichtung nach Anspruch 4, ferner eine Einrichtung zum Bilden eines Durchschnitts aus den kombinierten gewichteten ungespreizten Signalen aufweisend.

6. Vorrichtung nach Anspruch 5, ferner eine Einrichtung zum Wiederholen der Akkumulierungs-, Gewichtungs-, Kombinierungs- und Durchschnittsbildungsschritte während jedes Beobachtungszeitraums, in dem der Schwellenwert verwendet wird, aufweisend.

Revendications

1. Un procédé de calcul d'un seuil utilisé pour déterminer si une acquisition de code de pseudo-bruit (PN) s'est produite, le procédé comprenant les opérations suivantes :

l'accumulation de signaux reçus qui n'ont pas été désétalés à partir de chaque élément d'une pluralité de trajets reçus de façon à former une pluralité de signaux accumulés non étalés entretenus séparément,
la pondération de chaque signal accumulé non étalé provenant de chaque trajet reçu par une pondération correspondante à partir d'un vecteur de pondération, et
la combinaison des signaux accumulés non étalés pondérés.

2. Un procédé selon la Revendication 1, comprenant en outre l'opération de calcul de moyenne des signaux accumulés non étalés pondérés combinés.

3. Un procédé selon la Revendication 2, comprenant en outre l'opération de répétition des opérations d'accumulation, de pondération, de combinaison et de calcul de moyenne au cours de chaque période d'observation dans laquelle le seuil est utilisé.

4. Un appareil de calcul d'un seuil utilisé pour déterminer si une acquisition de code de pseudo-bruit (PN) s'est produite, l'appareil comprenant :

une pluralité de deuxièmes accumulateurs, chacun des deuxièmes accumulateurs étant raccordé de façon à recevoir un carré d'une sortie provenant d'un échantillonneur de puce correspondant, l'échantillonneur de puce étant raccordé de façon à recevoir un signal provenant d'un élément correspondant d'une pluralité de trajets reçus,
une pluralité de multiplicateurs, chaque multiplicateur étant raccordé à un accumulateur respectif de façon à recevoir une valeur accumulée finale, chaque multiplicateur étant configuré en outre de façon à pondérer la valeur accumulée finale par une pondération basée sur une pondération correspondante provenant d'un vecteur de pondération, et
un additionneur, l'additionneur étant raccordé à chacun des multiplicateurs de façon à former une somme des sorties provenant des multiplicateurs.

5. Un appareil selon la Revendication 4, comprenant en outre un moyen de calcul de moyenne des signaux accumulés non étalés pondérés combinés.

6. Un appareil selon la Revendication 5, comprenant en outre un moyen de répétition des opérations d'accumulation, de pondération, de combinaison et de calcul de moyenne au cours de chaque période d'observation dans laquelle le seuil est utilisé.

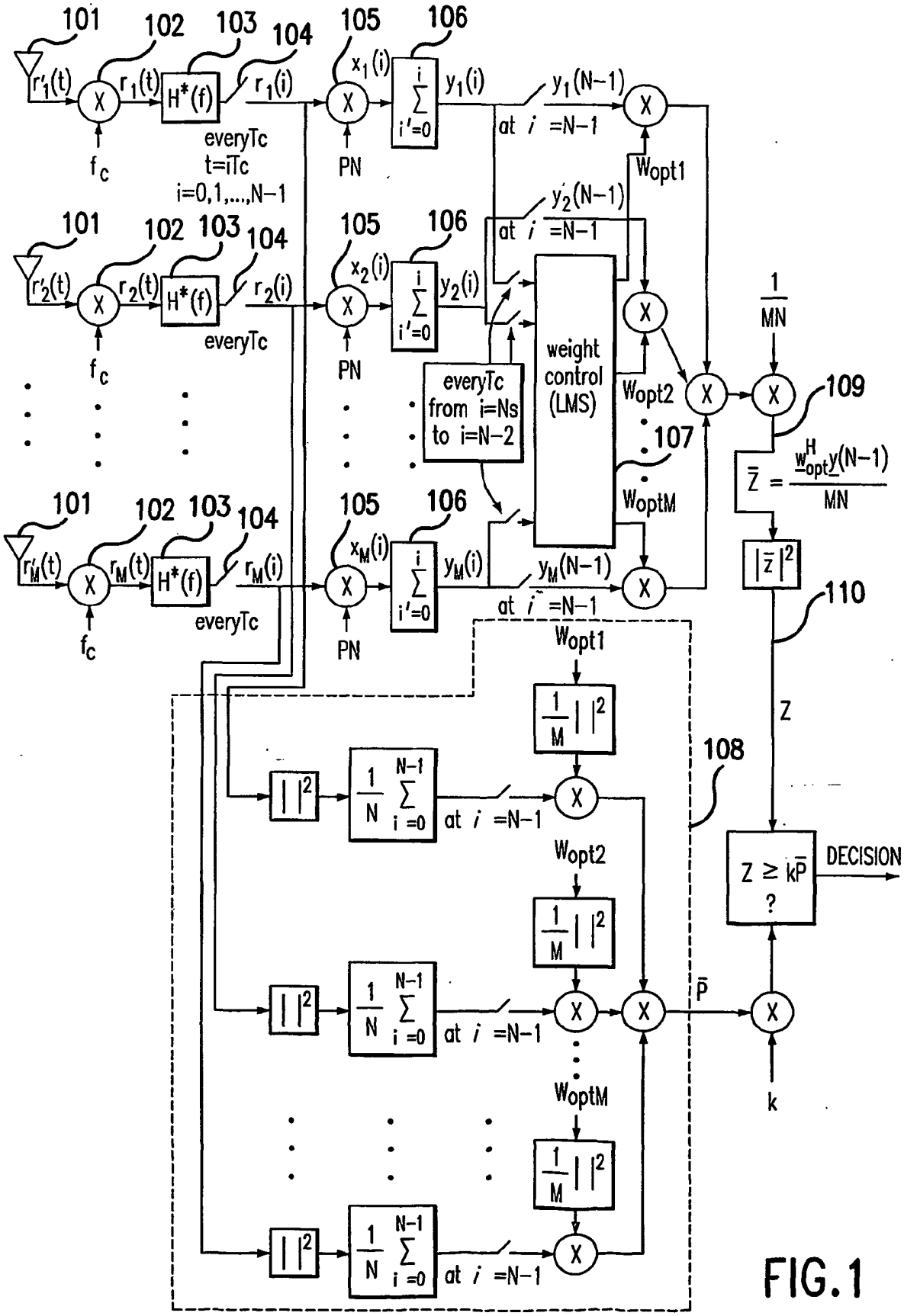
101
102
103
104
105
106
$r'_1(t)$
$r_1(t)$
$H^*(f)$
$r_1(i)$
$x_1(i)$
$y_1(i)$
$y_1(N-1)$
at i =N−1
f_c
everyTc
t=iTc
i=0,1,...,N−1
PN
W_{opt1}
$r'_2(t)$
$r_2(t)$
$r_2(i)$
$x_2(i)$
$y_2(i)$
$y'_2(N-1)$
everyTc
from i=Ns
to i=N−2
weight
control
(LMS)
W_{opt2}
W_{optM}
107
$\frac{1}{MN}$
109
$\bar{Z} = \frac{w_{opt}^H y(N-1)}{MN}$
$r'_M(t)$
$r_M(t)$
$r_M(i)$
$x_M(i)$
$y_M(i)$
$y_M(N-1)$
$|\bar{z}|^2$
110
Z
108
$\frac{1}{M}||^2$
$||^2$
$\frac{1}{N}\sum_{i=0}^{N-1}$
$\bar{P}$
$Z \geq k\bar{P}$
?
DECISION
k

FIG.1

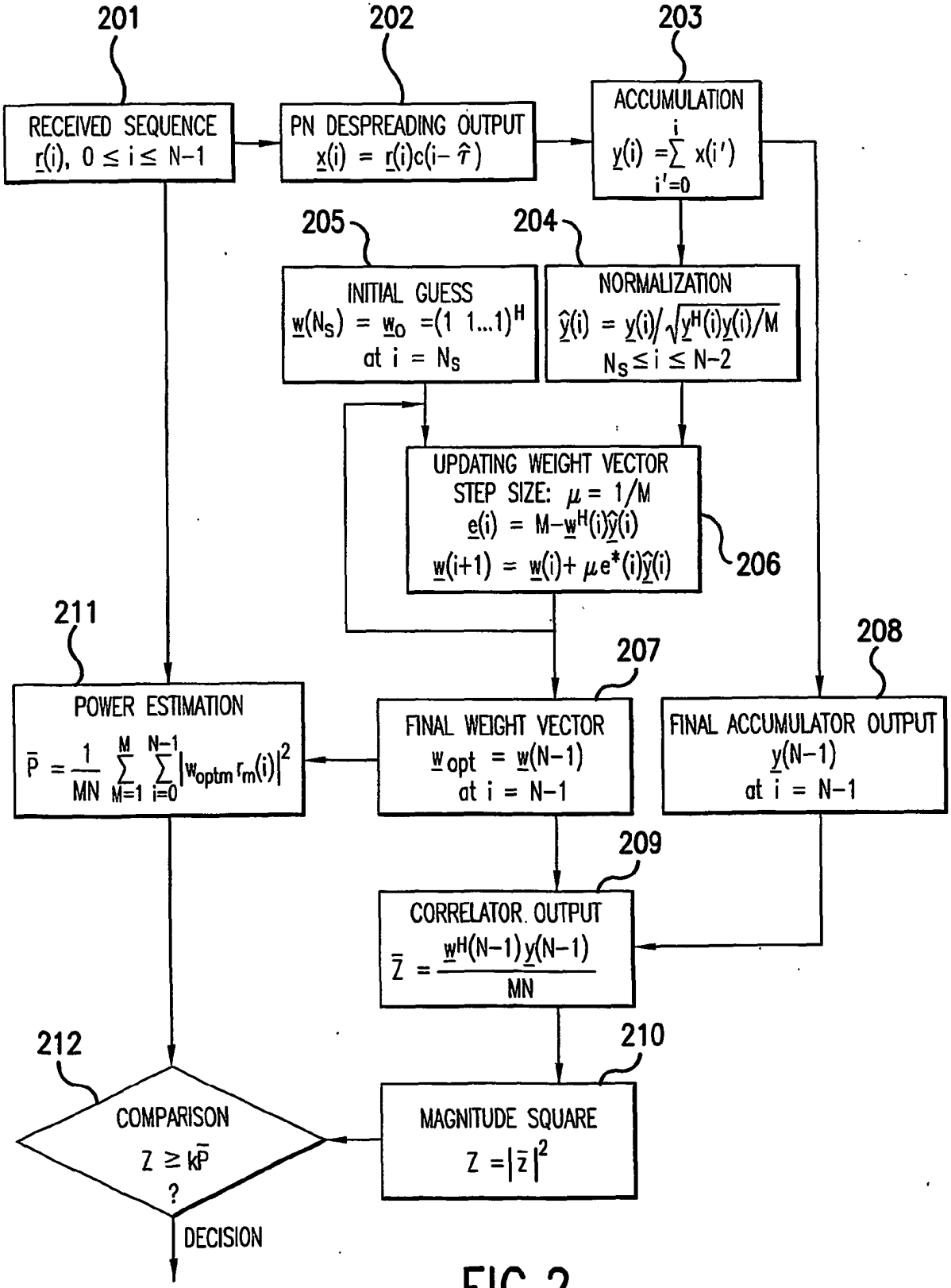
201
RECEIVED SEQUENCE
r(i), 0 ≤ i ≤ N-1
202
PN DESPREADING OUTPUT
x(i) = r(i)c(i- τ̂)
203
ACCUMULATION
y(i) = Σ x(i')
i'=0
205
INITIAL GUESS
w(N_S) = w_0 =(1 1...1)^H
at i = N_S
204
NORMALIZATION
ŷ(i) = y(i)/√(y^H(i)y(i)/M
N_S ≤ i ≤ N-2
UPDATING WEIGHT VECTOR
STEP SIZE: μ = 1/M
e(i) = M-w^H(i)ŷ(i)
w(i+1) = w(i)+ μe*(i)ŷ(i)
206
211
POWER ESTIMATION
P̄ = 1/MN Σ_{M=1}^{M} Σ_{i=0}^{N-1} |w_optm r_m(i)|^2
207
FINAL WEIGHT VECTOR
w_opt = w(N-1)
at i = N-1
208
FINAL ACCUMULATOR OUTPUT
y(N-1)
at i = N-1
209
CORRELATOR OUTPUT
Z̄ = w^H(N-1)y(N-1)/MN
210
MAGNITUDE SQUARE
Z = |z̄|^2
212
COMPARISON
Z ≥ kP̄
?
DECISION

FIG.2

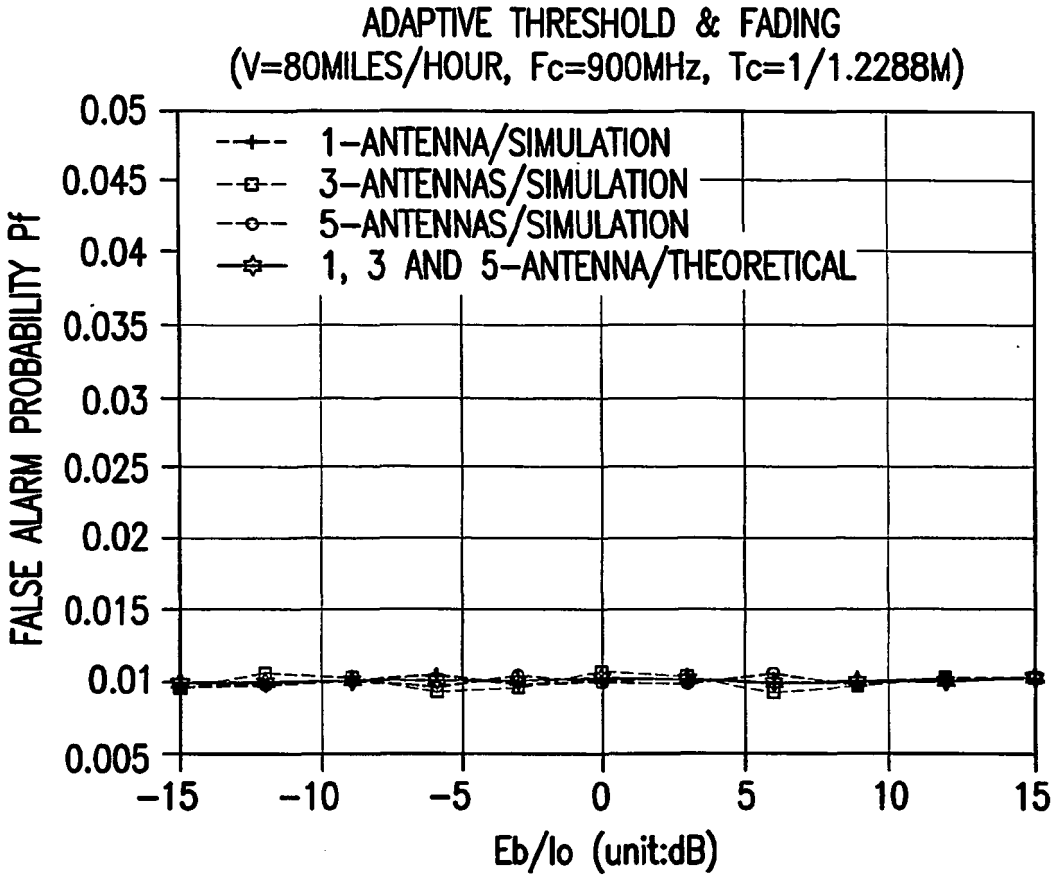
ADAPTIVE THRESHOLD & FADING
(V=80MILES/HOUR, Fc=900MHz, Tc=1/1.2288M)
1-ANTENNA/SIMULATION
3-ANTENNAS/SIMULATION
5-ANTENNAS/SIMULATION
1, 3 AND 5-ANTENNA/THEORETICAL
FALSE ALARM PROBABILITY Pf
0.05
0.045
0.04
0.035
0.03
0.025
0.02
0.015
0.01
0.005
-15
-10
-5
0
5
10
15
Eb/Io (unit:dB)

FIG.3

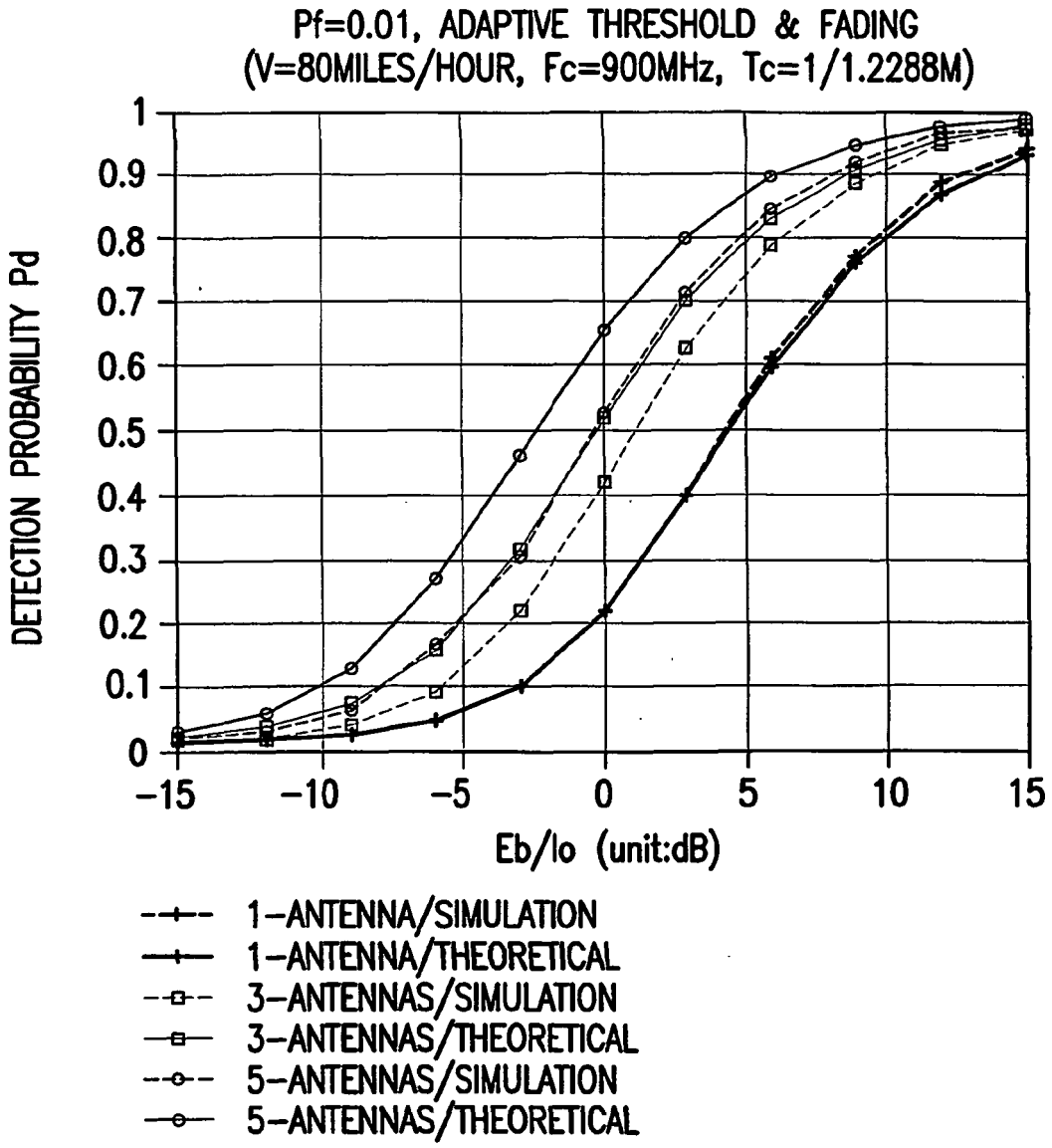
Pf=0.01, ADAPTIVE THRESHOLD & FADING
(V=80MILES/HOUR, Fc=900MHz, Tc=1/1.2288M)
DETECTION PROBABILITY Pd
1
0.9
0.8
0.7
0.6
0.5
0.4
0.3
0.2
0.1
0
−15
−10
−5
0
5
10
15
Eb/Io (unit:dB)
1-ANTENNA/SIMULATION
1-ANTENNA/THEORETICAL
3-ANTENNAS/SIMULATION
3-ANTENNAS/THEORETICAL
5-ANTENNAS/SIMULATION
5-ANTENNAS/THEORETICAL

FIG.4

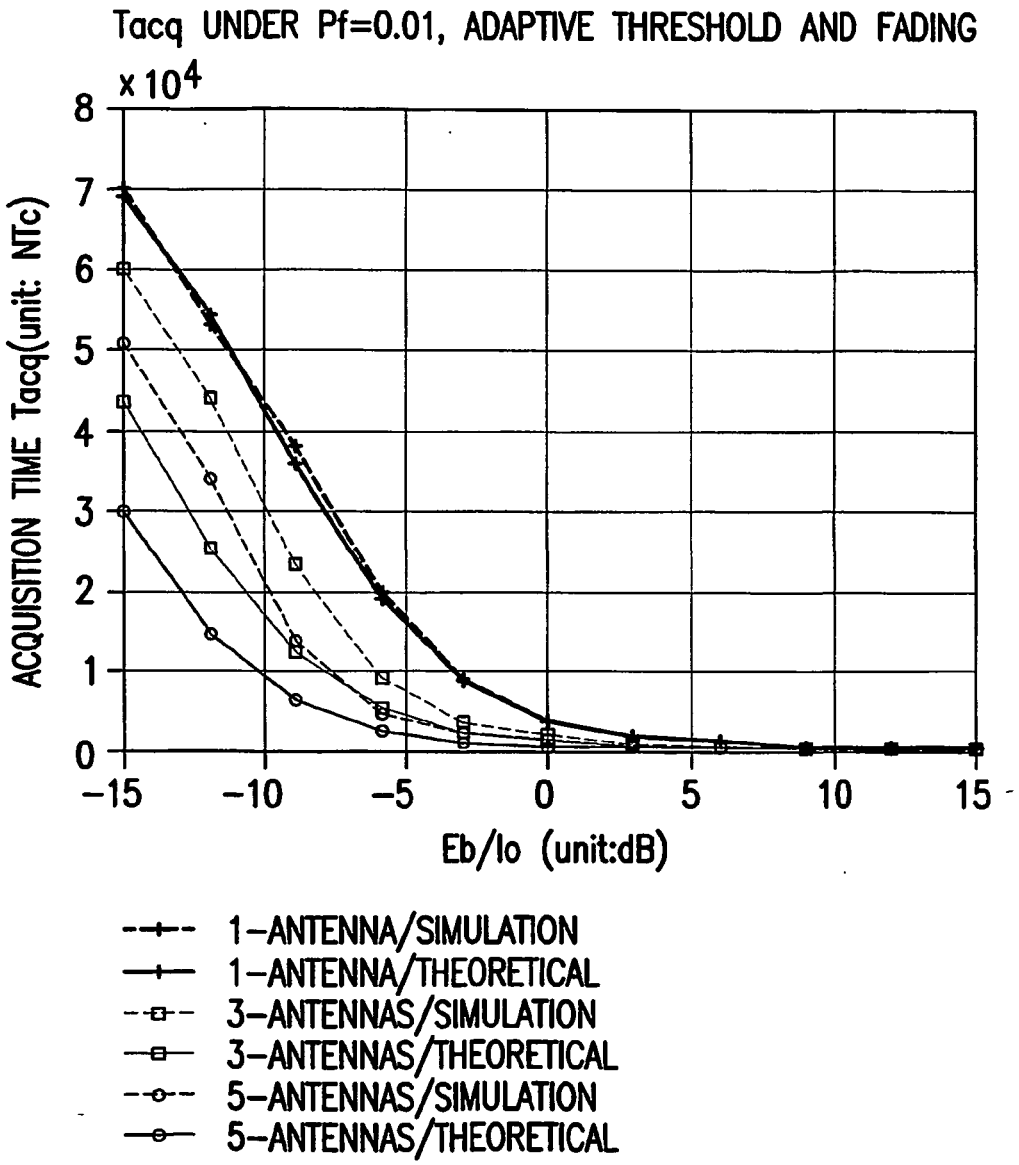
Tacq UNDER Pf=0.01, ADAPTIVE THRESHOLD AND FADING
x 10^4
ACQUISITION TIME Tacq(unit: NTc)
8
7
6
5
4
3
2
1
0
-15
-10
-5
0
5
10
15
Eb/Io (unit:dB)
1-ANTENNA/SIMULATION
1-ANTENNA/THEORETICAL
3-ANTENNAS/SIMULATION
3-ANTENNAS/THEORETICAL
5-ANTENNAS/SIMULATION
5-ANTENNAS/THEORETICAL

FIG.5

REFERENCES CITED IN THE DESCRIPTION

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